

## Correlation-based phase noise measurements

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(Received 7 March 2000; accepted for publication 26 April 2000)

In the characterization of the phase noise of a component, it is common practice to measure the cross-spectrum density at the output of two phase detectors that simultaneously compare the component output signal to a common reference. This technique, which is based on correlation and averaging, allows the rejection of the phase detector noise. On the other hand, it is known that the interferometer exhibits lower noise floor and higher conversion gain than other phase detectors suitable to radio-frequency and microwave bands. Thus, we experimented on an improved instrument in which the phase noise of a component is measured by correlating and averaging the output of two interferometers. The measurement sensitivity, given in terms of noise floor, turns out to be limited by the temperature uniformity of the instrument, instead of the absolute temperature  $T$ . This feature makes the instrument suitable to investigate the spectrum  $S_\varphi(f)$  of phase fluctuations below  $k_B T/P_o$ , i.e., the thermal energy  $k_B T$  referred to the carrier power  $P_o$ . The described method is suitable to the implementation of instruments in a wide frequency range, from some 100 kHz to 40 GHz and beyond. In principle, this method can also be exploited for the measurement of amplitude noise. Theory and experimental proof are given. © 2000 American Institute of Physics. [S0034-6748(00)01708-1]

### I. INTRODUCTION

For several reasons, some of which briefly considered underneath, phase noise plays a privileged role with respect to amplitude noise. In electronics, for instance, phase noise is generally integrated over a long time, while amplitude noise produces local effects only. Then, digital circuits are relatively immune to amplitude noise but prone to transition jitter, which results from phase noise. Furthermore, the Barkhausen conditions for steady oscillation (unity gain and zero phase) impose that the internal phase fluctuations of an oscillator are transformed into frequency fluctuations; this results in enhanced phase fluctuations at the output of the oscillator.<sup>1</sup> Finally, the frequency multiplication process also multiplies phase noise, which limits the maximum frequency attainable by frequency synthesis.

Phase noise is usually described in terms of the power spectrum density  $S_\varphi(f)$  of the phase fluctuation  $\varphi(t)$ . In radio frequency and microwave domains, up to 40 GHz or more, it is a common practice to measure  $S_\varphi(f)$  by means of a fast Fourier transform (FFT) analyzer preceded by a phase-to-voltage converter consisting of a saturated mixer. In good experimental conditions and with relatively high carrier level, some 15 dBm, the sensitivity of a noise measurement system based on a saturated mixer can hardly be higher than  $-170$  dBrad<sup>2</sup>/Hz (white noise) plus  $-140$  dBrad<sup>2</sup>/Hz at  $f = 1$  Hz (flicker). Higher sensitivity can be obtained by replacing the mixer with a radio-frequency interferometer,

which shows lower noise and higher conversion gain. In this case, sensitivity is limited by the absolute temperature of the interferometer and by the noise figure of an amplifier, and therefore, it is proportional to the carrier power  $P_o$ . For reference, with  $P_o = 10$  dBm the noise floor can approach  $-180$  dBrad<sup>2</sup>/Hz.

The interferometric method was initially proposed as a means to characterize microwave amplifiers.<sup>2</sup> Afterwards, it was used in conjunction with a discriminator for the measurement of the frequency stability of oscillators.<sup>3</sup> More recently, this method was exploited for the measurement of the phase noise of X-band passive devices<sup>4</sup> and for the frequency stabilization of whispering gallery oscillators accomplished by dynamical phase noise correction.<sup>5</sup> Then, a comprehension improvement<sup>6</sup> provided new design rules and the extension of the interferometric method to lower frequencies (100 MHz). Finally, the interferometer proved to be suitable to the measurement of the frequency stability of 5–10 MHz quartz resonators<sup>7</sup> with a sensitivity of some  $10^{-14}$ .

On the other hand, the sensitivity of phase noise measurements can be improved by exploiting a correlation technique, in which two equal instruments simultaneously measure the same device. If the two instruments are independent and only the device being tested is shared, the instrument noise is rejected. In frequency metrology, this technique was initially used as a means to measure the frequency stability of a hydrogen maser pair,<sup>8</sup> and subsequently repropounded in various ways,<sup>9–12</sup> with double-balanced mixers as the phase detector.

Thus, we combined the above two ideas, interferometer

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and correlation, proposing the double interferometer and implementing two prototypes.<sup>13,14</sup>

The noise floor of the double interferometer turns out to be limited by the temperature difference of the resistive terminations, instead of the absolute temperature. This noise compensation mechanism is similar to that of correlation radiometers and radio telescopes. But, our instrument is designed to measure the power spectrum density of noise close to a strong carrier signal, say 0–20 dBm, instead of the power of small radiation alone. A noise floor of some –195 dBrad<sup>2</sup>/Hz can be attained with a carrier power below 10 dBm.

The double interferometer, although originally intended as a means to characterize components for high stability oscillators, is actually a tool for general experiments involving the measurement of low-noise phenomena in the vicinity of a strong carrier signal. In addition, as the double interferometer removes the thermal floor, it makes possible the measurement of flicker noise at higher Fourier frequencies than other instruments.

## II. BASIC CONCEPTS

A high signal-to-noise ratio sinusoidal signal  $s(t)$  of frequency  $\nu_0$  and power  $P_o$  at the output of a source impedance matched to its characteristic impedance  $R_0$  can be represented as

$$s(t) = \sqrt{2R_0P_o} [1 + \alpha(t)] \sin[2\pi\nu_0 t + \varphi(t)]. \quad (1)$$

$\varphi(t)$  and  $\alpha(t)$  are the phase modulation (PM) noise and the amplitude modulation (AM) noise, respectively. The physical quantity of major interest is the power spectrum density (PSD)  $S_\varphi(f)$  of  $\varphi(t)$  as a function of the Fourier frequency  $f$ .

Signal (1) can be rewritten as

$$s(t) = \sqrt{2R_0P_o} \sin[2\pi\nu_0 t] + n(t), \quad (2)$$

where  $n(t)$  is the random voltage that causes AM and PM noise. Then,  $n(t)$  can be divided as

$$n(t) = n_c(t) \cos(2\pi\nu_0 t) + n_s(t) \sin(2\pi\nu_0 t), \quad (3)$$

which is related to the AM and PM noise by

$$\alpha(t) = \frac{n_s(t)}{\sqrt{R_0P_o}} \quad \text{and} \quad \varphi(t) = \frac{n_c(t)}{\sqrt{R_0P_o}}. \quad (4)$$

The spectrum densities  $N_c(f)$  and  $N_s(f)$  come from the superposition of the upper and lower sidebands of  $n(t)$ , i.e.,  $N(\nu_0 + f)$  and  $N(\nu_0 - f)$ , which is inherent in the frequency conversion process.

If  $n(t)$  is a true additive random voltage, like thermal noise is,  $n_c(t)$  and  $n_s(t)$  are independent random variables of equal PSD. We assume that even if  $n(t)$  is of parametric origin,  $n_c(t)$  and  $n_s(t)$  are independent, although their PSDs may be different. In most cases, this approximation is close to the actual behavior of radiofrequency and microwave devices. In fact, in the presence of a carrier signal the noise phenomena of these devices tend to affect phase and amplitude independently. Moreover,  $n(t)$  can be represented as  $n(t) = n^{\text{th}}(t) + n^{\text{ex}}(t)$ , which is the superposition of thermal noise  $n^{\text{th}}(t)$  and extra noise  $n^{\text{ex}}(t)$ ; we use the word “extra”

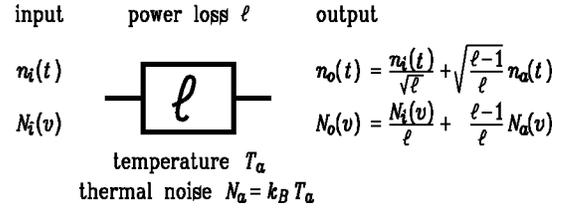


FIG. 1. Noise model of an attenuator.

to avoid “excess” because the latter is often considered synonymous of flicker, which is more restrictive. Obviously, AM and PM noise, as well as spectra, can be divided in the same way. Furthermore, we assume that  $n^{\text{th}}(t)$  and  $n^{\text{ex}}(t)$  are independent. Finally,  $n^{\text{th}}(t)$  and  $n^{\text{ex}}(t)$  can be separately decomposed according to Eq. (3).

As a consequence of the Nyquist theorem, a resistor of value  $R_0$  at temperature  $T$  can be modeled with a cold resistor  $R_0$  in series to a generator of random voltage  $2n(t)$  that accounts for thermal noise. Hence, a voltage  $n(t)$  is available across an impedance-matched load. The corresponding PSD is  $N(\nu) = k_B T R_0$ , where  $k_B = 1.38 \times 10^{-23}$  is the Boltzmann constant. With thermal noise, it holds  $N_c(f) = k_B T R_0$  and  $N_s(f) = k_B T R_0$ . Thus, a sinusoidal signal from a source whose internal resistance is at temperature  $T$  is affected by AM noise  $S_\alpha^{\text{th}}(f) = k_B T / P_o$  and by PM noise  $S_\varphi^{\text{th}}(f) = k_B T / P_o$ .

The extension of the Nyquist theorem for circuits at non-uniform temperature yields the description of the attenuator behavior shown in Fig. 1. The attenuator, of power-loss  $l$ , is impedance matched to  $R_0$  at both ends, and is at temperature  $T_a$ . When the thermal noise  $n_i(t)$  from a resistance  $R_0$  at temperature  $T_i$  crosses this attenuator, it results in a random signal  $n'_o(t) = n_i(t) / \sqrt{l}$  at the output. The attenuator adds its noise. Indicating with  $n_a(t)$  the equivalent noise voltage of a resistance  $R_0$  at the temperature  $T_a$ , the noise contribution of the attenuator is  $n''_o(t) = \sqrt{(1-l)/l} n_a(t)$ . This is related to the fact that if the attenuator and the resistor are at the same temperature  $T$  the total output spectrum density  $N_o(\nu) = N'_o(\nu) + N''_o(\nu)$  must be equal to  $k_B T R_0$  and independent of  $l$ , and that for  $l = \infty$  the equivalent temperature observed at the output is  $T_a$ .

Let us now consider the sum and the difference of two independent random signals  $n_1(t)$  and  $n_2(t)$ :

$$a(t) = \frac{1}{\sqrt{2}} [n_1(t) + n_2(t)], \quad (5)$$

$$b(t) = \frac{1}{\sqrt{2}} [n_1(t) - n_2(t)], \quad (6)$$

obtained either as the result of a mere algebraic operation or by means of a lossless 3 dB coupler; such a coupler, which does not add thermal noise, is the idealization of a 3 dB hybrid junction, i.e., a transformer, a microstrip network, or a microwave magic  $T$ . As we need not a representation of form (1) or (3), we can use the baseband frequency  $f$ , which is consistent with the notation of Sec. IV. The cross PSD of  $a(t)$  and  $b(t)$  is, by definition,

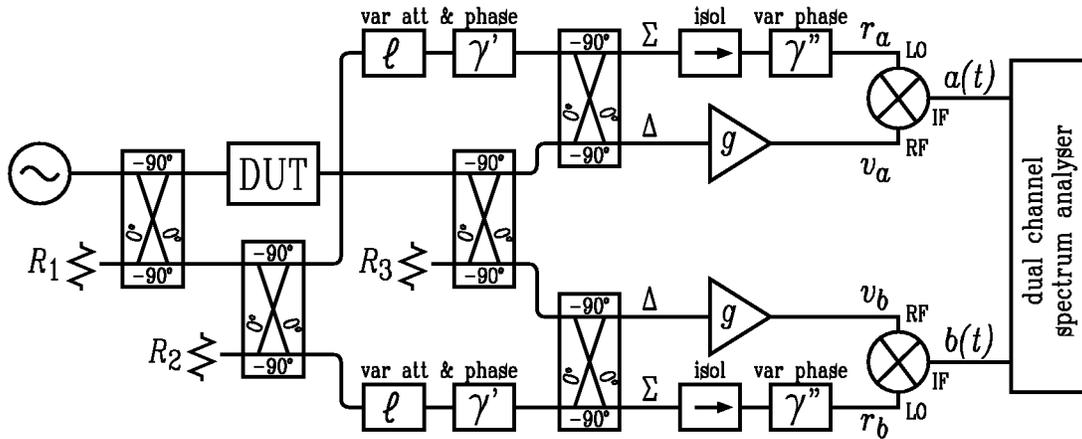


FIG. 2. Scheme of the double interferometer.

$$S_{ab}(f) = \mathcal{F}\{\mathcal{R}_{ab}(\tau)\} = \int_{-\infty}^{\infty} \mathcal{R}_{ab}(\tau) \exp(-2\pi f\tau) d\tau, \quad (7)$$

where  $\mathcal{F}\{\cdot\}$  is the Fourier transform operator, and  $\mathcal{R}_{ab}(\tau)$  is the cross-correlation function

$$\mathcal{R}_{ab}(\tau) = \lim_{\theta \rightarrow \infty} \frac{1}{\theta} \int_{\theta}^{\theta + \tau} a(t) b^*(t - \tau) dt; \quad (8)$$

the symbol “\*” stands for complex conjugate and can be omitted because we deal with real signals. Making  $n_1(t)$  and  $n_2(t)$  appear in Eq. (7), and dividing thermal and extra noise, we get

$$S_{ab}(f) = \frac{1}{2} [N_1(f) - N_2(f)] \quad (9)$$

$$= \frac{1}{2} [k_B(T_1 - T_2)R_0 + N_1^{\text{ex}}(f) - N_2^{\text{ex}}(f)]. \quad (10)$$

This means that if  $n_2(t)$  is a pure thermal fluctuation and the temperature of the instrument is homogeneous ( $T_1 = T_2$ ), the instrument compensates for thermal noise and measures the extra noise  $N_1^{\text{ex}}(f)$  only. Alternatively, in the absence of extra noise, the instrument noise floor is limited by the temperature inhomogeneity  $T_1 - T_2$ .

$S_{ab}(f)$  is related to the Fourier transform  $A(f)$  and  $B(f)$  of the individual signals by

$$S_{ab}(f) = A(f) B^*(f). \quad (11)$$

Generally, dynamic signal analyzers make use of Eq. (11) replacing the true Fourier transform with the FFT of  $a(t)$  and  $b(t)$  simultaneously sampled, and averaging over  $m$  acquisitions. Thus, the estimate of the spectrum  $S_{ab}(f)$  is affected by a rms uncertainty

$$\sigma_{S_{ab}} = \frac{|A| |B|}{\sqrt{2m}}. \quad (12)$$

If  $a(t)$  and  $b(t)$  are uncorrelated, the estimate of  $S_{ab}(f)$  approaches zero proportionally to  $1/\sqrt{2m}$ , being limited by the uncertainty (12). Therefore, a long averaging time may be needed to attain the ultimate noise floor of the instrument, determined by temperature inhomogeneity or cross talk.

It should be remarked that the high sensitivity of the correlation microwave radio telescopes<sup>15,16</sup> relies upon Eq.

(10). A similar mechanism is exploited in the Allred radiometer<sup>17,18</sup> to compare a noise source to a reference one. This instrument works as a sort of bridge that is nulled observing the sign of  $S_{ab}$  and controlling the reference source. The noise compensation mechanism (10), combined with a carrier suppression technique and down conversion, makes the realization of our high-sensitivity phase noise measurements system possible.

### III. PHASE NOISE MEASUREMENT SYSTEM

The theory of the interferometer is reported in Ref. 6, together with design strategies and experimental results. The double interferometer, shown in Fig. 2, consists of two interferometers that simultaneously measure the phase noise of a single device under test (DUT). Setting the variable attenuators  $l$  and the variable phase shifters  $\gamma'$  equal to the DUT phase and attenuation, all the oscillator power goes to the  $\Sigma$  port of the hybrids, while the carrier is suppressed at the  $\Delta$  outputs. The carrier suppression mechanism has no effect on the DUT noise. Therefore, one fourth of the power of the DUT noise sidebands is present at the input of each amplifier. Setting the phases  $\gamma''$  equal to the phase lag of the amplifiers, the mixers down convert the DUT phase noise to the baseband. In this condition, the voltages  $a(t)$  and  $b(t)$  present at the output of the mixers are proportional to the instant value of the DUT phase  $\varphi(t)$ . Consequently, the cross PSD of the two output signals is proportional to the DUT phase noise PSD, while the individual interferometer noise is rejected.

The three hybrids on the left part of Fig. 2 are used as power splitters and may be replaced with them. It should be remarked that the power splitter is actually a four-port hybrid internally terminated at one port, otherwise it could not be impedance matched at all ports. The scheme of Fig. 2 is based on 90° hybrids, which corresponds to the 100 MHz implementation used in this article. Yet, any other combination of 180° and 90° hybrids and power splitters would work in the same way. The only constraint is that the carrier suppression and the detection of the DUT phase noise must be ensured by properly setting  $\gamma'$  and  $\gamma''$ .

#### IV. SIGNAL ANALYSIS OF THE DOUBLE INTERFEROMETER

Let us now analyze in detail the double-interferometer setup for the measurement of a generic DUT of loss  $l$  that produces both thermal and extra noise.

As the oscillator provides the phase reference to the whole machine, its phase noise is rejected. The noise of the variable attenuators and phase shifters responsible for the carrier suppression vanishes in the correlation and averaging process because these devices are independent and the two arms are isolated. The noise of the two amplifiers vanishes for the same reason. The mixers and the detection phase shifters  $\gamma''$  are independent and they process amplified signals, thus their noise contribution is negligible. At a deeper sight, one can observe that the effect of all the independent noise sources (attenuators, phase shifters, amplifiers, etc.) can only be an increase in the number  $m$  of averages needed for a given noise floor, according to Eq. (12). For the sake of simplicity, we assume that the hybrids are lossless and noiseless; anyway, nonideality of the hybrids can be reintroduced later in the equations.

Under the above assumptions, there remain four random signals shared by the two arms, namely, the DUT noise  $n_d(t)$  and the thermal noise  $n_1(t)$ ,  $n_2(t)$ , and  $n_3(t)$  of the resistive terminations  $R_1$ ,  $R_2$ , and  $R_3$  at temperature  $T_1$ ,  $T_2$ , and  $T_3$ ; as there is no ambiguity, we omit the superscript ‘‘th’’ of  $n_1(t)$ ,  $n_2(t)$ , and  $n_3(t)$ . Defining the thermal noise of a resistor at temperature  $T_d$  as  $n_d^{\text{th}}(t)$ , the model of Fig. 1 yields a DUT noise contribution of  $\sqrt{(l-1)/l} n_d^{\text{th}}(t)$ . Besides this, we define the extra noise at the DUT output as  $n_d^{\text{ex}}(t)$ .

When the oscillator signal  $\sqrt{2R_0P_r} \cos(2\pi\nu_0t)$  is taken as the phase reference, the phase of the DUT output signal is that of  $\sin(2\pi\nu_0t)$ . Hence, phase noise comes from the  $n_c(t)\cos(2\pi\nu_0t)$  component of the DUT noise. Consequently, it holds  $S_\varphi(f) = N_c(f)/(R_0P_o)$ , where  $P_o$  is the carrier power at the DUT output.

The reference signals at the mixer LO ports are

$$r_a(t) = -V_p \cos(2\pi\nu_0t), \quad (13)$$

$$r_b(t) = V_p \sin(2\pi\nu_0t). \quad (14)$$

Accordingly, arm  $a$  detects the  $\cos(2\pi\nu_0t)$  component of the signal present at the rf port of the mixer, and arm  $b$  detects the  $\sin(2\pi\nu_0t)$  component. These rf signals are

$$\begin{aligned} v_a(t) = & \sqrt{g} \left[ -\frac{1}{\sqrt{2l}} n_{1c}(t) + \frac{1}{2\sqrt{l}} n_{2c}(t) + \frac{1}{2} n_{3s}(t) \right. \\ & \left. - \frac{1}{2} \sqrt{\frac{l-1}{l}} n_{dc}^{\text{th}}(t) - \frac{1}{2} n_{dc}^{\text{ex}}(t) \right] \cos(2\pi\nu_0t) \\ & + \sqrt{g} \begin{bmatrix} \text{NON} \\ \text{DETECTED} \\ \text{TERMS} \end{bmatrix} \sin(2\pi\nu_0t), \end{aligned} \quad (15)$$

$$\begin{aligned} v_b(t) = & \sqrt{g} \left[ \frac{1}{\sqrt{2l}} n_{1c}(t) + \frac{1}{2\sqrt{l}} n_{2c}(t) + \frac{1}{2} n_{3s}(t) \right. \\ & \left. + \frac{1}{2} \sqrt{\frac{l-1}{l}} n_{dc}^{\text{th}}(t) + \frac{1}{2} n_{dc}^{\text{ex}}(t) \right] \sin(2\pi\nu_0t) \\ & + \sqrt{g} \begin{bmatrix} \text{NON} \\ \text{DETECTED} \\ \text{TERMS} \end{bmatrix} \cos(2\pi\nu_0t). \end{aligned} \quad (16)$$

After filtering out the  $2\nu_0$  components, the down-converted signals are

$$\begin{aligned} a(t) = & \sqrt{\frac{2g}{l_m}} \left[ \frac{1}{\sqrt{2l}} n_{1c}(t) - \frac{1}{2\sqrt{l}} n_{2c}(t) - \frac{1}{2} n_{3s}(t) \right. \\ & \left. + \frac{1}{2} \sqrt{\frac{l-1}{l}} n_{dc}^{\text{th}}(t) + \frac{1}{2} n_{dc}^{\text{ex}}(t) \right], \end{aligned} \quad (17)$$

$$\begin{aligned} b(t) = & \sqrt{\frac{2g}{l_m}} \left[ \frac{1}{\sqrt{2l}} n_{1c}(t) + \frac{1}{2\sqrt{l}} n_{2c}(t) + \frac{1}{2} n_{3s}(t) \right. \\ & \left. + \frac{1}{2} \sqrt{\frac{l-1}{l}} n_{dc}^{\text{th}}(t) + \frac{1}{2} n_{dc}^{\text{ex}}(t) \right], \end{aligned} \quad (18)$$

where  $l_m$  is the mixer loss; according to the usual definition,  $l_m$  includes the 3 dB intrinsic loss due to the fact that the mixer makes the sum and the difference of its input frequencies, and consequently, it splits the input power into two bands.

The cross correlation  $\mathcal{R}_{ab}(\tau)$  contains only the autocorrelation terms, while the mixed ones vanish because all the noise processes are independent. Hence, the cross PSD is

$$\begin{aligned} S_{ab}(f) = & \frac{g}{l_m} \left[ \frac{1}{l} N_{1c}(f) - \frac{1}{2l} N_{2c}(f) - \frac{1}{2} N_{3s}(f) \right. \\ & \left. + \frac{l-1}{2l} N_{dc}^{\text{th}}(f) + \frac{1}{2} N_{dc}^{\text{ex}}(f) \right], \end{aligned} \quad (19)$$

and therefore,

$$\begin{aligned} S_{ab}(f) = & \frac{gk_B}{l_m} \left[ \frac{1}{l} T_1 - \frac{1}{2l} T_2 - \frac{1}{2} T_3 + \frac{l-1}{2l} T_d \right] \\ & + \frac{g}{2l_m} N_{dc}^{\text{ex}}(f). \end{aligned} \quad (20)$$

If the whole machine is at the same temperature  $T=T_0$ , the cross PSD reduces to

$$S_{ab}(f) = \frac{g}{2l_m} N_{dc}^{\text{ex}}(f), \quad (21)$$

which is determined by the DUT extra noise only. Therefore, under the hypothesis of temperature uniformity, only the extra noise  $N_{dc}^{\text{ex}}(f)$  contributes to the observed spectrum. This means that the simple noise compensation mechanism given by Eq. (10) also takes place in the double interferometer, which makes the thermal noise vanish.

To derive the gain  $K_\varphi = S_{ab}(f)/S_\varphi(f)$  of the double interferometer we assume that thermal noise is negligible compared to the DUT extra noise, i.e.,  $N_d(f) \approx N_d^{ex}(f)$ . Then, inserting  $S_\varphi(f) = N_c(f)/(R_0 P_o)$  in Eq. (21) we get

$$K_\varphi = \frac{gR_0P_o}{2l_m}. \quad (22)$$

As there are two hybrids along the signal path from the DUT to the amplifier, each of which shows a loss  $l_h$ , actual gain is lower than Eq. (22) by a factor  $1/l_h^2$ .

The single-arm noise floor  $S_{\varphi_0}(f)$  can be derived from the equivalent noise  $Fk_B T_0$  at the amplifier input, where  $F$  is the amplifier noise figure. The PSD of the down converted voltage, either  $a(t)$  or  $b(t)$ , is  $S_{v_0}(f) = 2Fgk_B T_0 R_0 / l_m$ . Dividing the latter by gain (22), we get

$$S_{\varphi_0}(f) = 4 \frac{Fk_B T_0}{R_0 P_o}. \quad (23)$$

Then, accounting for the hybrid loss, the actual noise floor is higher than Eq. (23) by a factor  $l_h^2$ .

Finally, the double interferometer can be set up to detect amplitude noise. This is easily accomplished by adding  $90^\circ$  to the two phase shifters  $\gamma''$ , which causes the  $n_s(t)$  component of the DUT noise to be down converted instead of the  $n_c(t)$  one. Equations (16)–(23) still hold, provided some obvious subscript changes were done.

## V. EXPERIMENTAL PROOF AND DISCUSSION

### A. Noise floor

We measured the noise floor of a double-interferometer prototype designed for the carrier frequency  $\nu_0 = 100$  MHz. To do so, the DUT is replaced with a short cable, which is noiseless. In this prototype, the hybrids show a loss  $l_h = 0.8$  dB, while the loss of the mixers is  $l_m = 6$  dB. The signal power at the mixer LO inputs is 8 dBm, and the DUT power is  $P_o = 8$  dBm. The amplifiers show a gain  $g = 40$  dB, a noise figure  $F = 2$  dB, and a bandwidth of some 30 MHz centered around  $\nu_0$ . All the circuits are impedance matched to  $R_0 = 50 \Omega$ . Further details of this instrument are reported in Ref. 14.

Properly adjusting the interferometers, the carrier suppression is at least 65 dB. In this condition, the amplifier linearity is ensured because the residual output carrier never exceeds  $-25$  dBm, which is 40 dB lower than the 1 dB compression point of the amplifiers.

In order to set the phases  $\gamma''$  for the two mixers to detect phase noise, the DUT is replaced with a small-angle phase modulator driven by the oscillator output of a lock-in. Thus, a sinusoidal modulation of the order of 1 mrad is injected. Then, each phase shifter is adjusted for zero voltage at the output of the corresponding mixer, which is measured with the lock-in. Finally, two  $90^\circ$  cables calibrated by means of a network analyzer are added to  $\gamma''$ . Due to the high sensitivity of the described null method, phase accuracy and phase matching of the order of  $1^\circ$  can easily be obtained. The phase modulator and the lock-in are also used to measure the gain  $K_\varphi$ , which is  $24.5 \text{ dBV}^2/\text{rad}^2$  in this case.

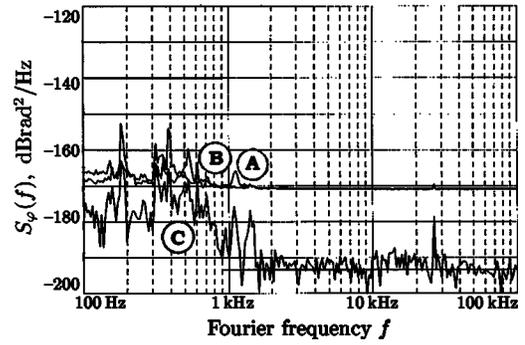


FIG. 3. Noise floor of the 100 MHz double-interferometer prototype. A and B: single arm. C: correlation.

The rejection of the driving oscillator noise turns out to be higher than 60 dB in the described conditions. Using a low-noise quartz oscillator that exhibits a floor of  $-157$  dBrad<sup>2</sup>/Hz, the noise contribution of that oscillator is negligible for our purposes.

Figure 3 shows the results averaged on  $m = 32767$  measures, which is the maximum of the available spectrum analyzer. Flicker and acoustic vibrations appear in the left part of the plot, while for  $f \geq 2$  kHz white noise only is present. The single-arm noise floor (curves A and B) is  $-172$  dBrad<sup>2</sup>/Hz, which is close to the expected value  $S_{\varphi_0}(f) = 4Fk_B T_0 l_h^2 / P_o \approx -172.3$  dBrad<sup>2</sup>/Hz. For reference, the thermal noise of a resistor at  $T_0 = 290$  K referred to the same carrier power  $P_o$  is  $S_\varphi^{th}(f) = k_B T_0 / P_o \approx -182$  dBrad<sup>2</sup>/Hz. Yet, the cross spectrum floor (curve C) is  $-194$  dBrad<sup>2</sup>/Hz, which is 12 dB lower than the thermal floor  $S_\varphi^{th}(f)$ .

As a consequence of Eq. (12), a rms uncertainty  $S_{\varphi_0} / \sqrt{2m} \approx -196$  dBrad<sup>2</sup>/Hz is expected, which is close to the observed floor. Consequently, that floor is due to the insufficient averaging capability of the analyzer instead of a true hardware limitation, and it is expected to further decrease, increasing  $m$ .

### B. Noise of an attenuator

The analytical development of Eq. (21) predicts that the thermal noise of an attenuator of loss  $l$  inserted as the DUT vanishes in the correlation-and-averaging process if the temperature of the whole machine is uniform, and that this is independent of  $l$ . We demonstrate this fact through an experiment in which two different configurations are compared. Thus, the DUT and the hybrid present in the center of Fig. 2 are replaced with a three port device that takes the two configurations of Fig. 4. In (A), most of the output noise comes from the two independent  $l = 16$  dB attenuators, while in (B) there is only one  $l = 16$  dB attenuator along the shared path. Yet, the cross PSD is expected to approach zero in both cases.

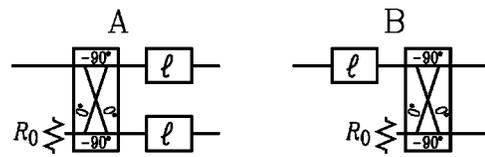


FIG. 4. Measurement schemes with the 16 dB attenuators.

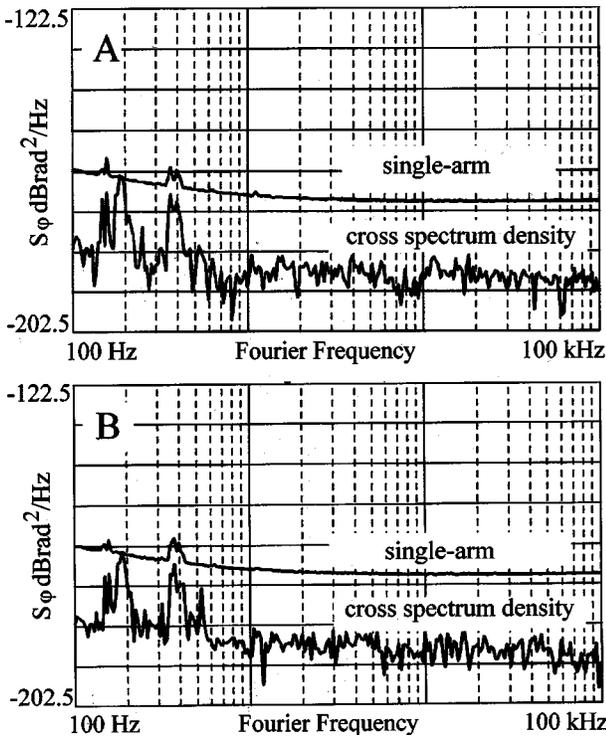


FIG. 5. Phase noise measured with the two 16 dB attenuator configurations.

All the experimental conditions are the same for the two configurations, except for the detail of Fig. 4. Thus, the hybrid driving power is different, but this is not relevant because the hybrid is not power sensitive. Moreover, we choose a device specified for continuous operation up to 36 dBm, while the input power does not exceed 22 dBm. The oscillator power is  $P_r = 26$  dBm. Accordingly, the output power is  $P_o = 6$  dBm at each port, the same for both configurations. All other circuit parameters, as well as the adjustment and calibration, are the same as described in Sec. V A.

The measured PSDs, averaged on  $m = 1024$  measures, are shown in Fig. 5. We focus our attention on the white noise floor, for  $f > 1$  kHz. According to Eq. (23), the expected single-arm noise is  $S_{\varphi_0}(f) = -170.4$  dBrad<sup>2</sup>/Hz, which is experimentally confirmed with both configurations.

The cross PSD floor is  $-187$  dBrad<sup>2</sup>/Hz, almost equal for both configurations. That floor is lower than the thermal energy referred to the carrier power, i.e.,  $S_{\varphi}^{\text{th}} = k_B T_0 / P_o = -182$  dBrad<sup>2</sup>/Hz, and corresponds to the uncertainty limit given by Eq. (12).

### C. Low-noise measurement

The DUT is now replaced with the circuit shown in Fig. 6, which enables the injection of extra noise  $N_d^{\text{ex}}(\nu)$

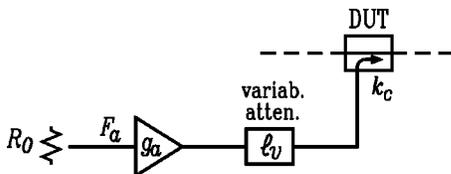


FIG. 6. Injection of the reference noise  $N_d^{\text{ex}}$ .

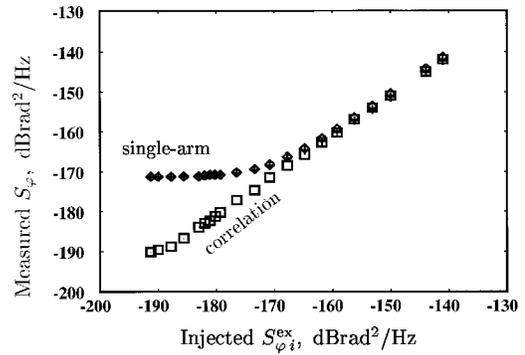


FIG. 7. Reference noise measurement.

$= g_a R_0 F_a k_B T_0 / (l_v k_c)$  through a directional coupler. This corresponds to a phase noise  $S_{\varphi_i}^{\text{ex}} = g_a R_0 F_a k_B T_0 / (l_v k_c P_o)$ , plus an amplitude noise of the same value; the latter is negligible because it is not detected. The source PSD is  $F_a k_B T_0 g_a \approx -100.5$  dBm/Hz, while the coupling factor is  $k_c = 11.5$  dB. The attenuator consists of a 0–70 dB variable unit in series to a 20 dB fixed one; for best impedance matching and stability, the fixed attenuator is located close to the directional coupler. As the DUT power is  $P_o = 8$  dBm, the injected extra noise can be set to the desired value in the  $-140$  to  $-210$  dBrad<sup>2</sup>/Hz range. In addition, the presence of thermal noise is ensured by impedance matching. The usefulness of this arrangement consists of the capability of calibrating a subthermal random signal by means of relatively large signal measurements only, which can be easily performed with conventional spectrum and network analyzers.

All the operating parameters of this experiment are the same as those reported in Sec. V A except for  $m$ , which is reduced when possible.

Figure 7 shows the measured floor  $S_{\varphi_0}$  averaged over  $f$  from some 5 to 50 kHz, as a function of the injected  $S_{\varphi_i}^{\text{ex}}$ . Going towards the left of Fig. 7,  $l_v$  increases and the injected noise becomes negligible compared to the single-arm noise floor. Therefore, the single-arm phase noise approaches the  $S_{\varphi_0} = -172$  dBrad<sup>2</sup>/Hz, the same value that was measured in the absence of the DUT. By contrast, the correlated noise fits the straight line  $S_{\varphi} = S_{\varphi_i}^{\text{ex}}$ . Consequently, still under the hypothesis of temperature uniformity, the instrument compensates for the thermal noise and measures the extra noise only. This confirms what we expect from Eq. (21).

### ACKNOWLEDGMENTS

The authors wish to acknowledge Michele Elia from the Politecnico di Torino for having greatly contributed to the theoretical comprehension of our experiments, and Marcel Olivier and Jacques Gros Lambert from the LPMO for their support and for valuable scientific discussions.

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